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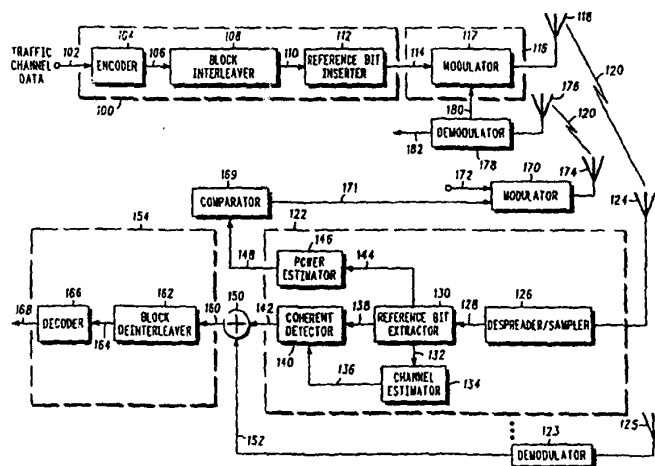
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(54) Title: METHOD AND APPARATUS FOR POWER ESTIMATION IN A COMMUNICATION SYSTEM



(57) Abstract

A method and apparatus is provided for estimating signal power. This is accomplished by demodulating a received communication signal to derive a stream of samples. These samples may consist of either reference samples (132), data samples (142), or a combination thereof. An estimate of the power (118) of the received communication signal is then generated as a function of the stream of samples (144). The signal power estimate (148) may, for example, be used in a power control algorithm by setting a power control indicator (121), in response to a comparison between the signal power estimate and a predetermined threshold. Subsequently, the power control indicator (171) is transmitted over a communication channel to another communication unit. The communication unit detects a power control indicator (171) within a signal (120) received from over the communication channel and adjusts a particular signal transmission power of a signal transmitter in response to the detected power control indicator (171).

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METHOD AND APPARATUS FOR POWER ESTIMATION IN A COMMUNICATION SYSTEM

Related Inventions

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The present invention is related to the following invention which is assigned to the assignee of the present invention. Method and Apparatus for Coherent Communication in a Spread-Spectrum Communication System by Fuyun Ling having U.S. Serial No. 10 08/031,258, and filed on March 11, 1993.

Field of the Invention

15 The present invention relates to communication systems and, more particularly, to a method and apparatus for signal power estimation in a communication system.

Background of the Invention

20

Communication systems take many forms. In general, the purpose of a communication system is to transmit information-bearing signals from a source, located at one point, to a user destination, located at another point some distance away. A communication system generally consists of three basic components: transmitter, channel, and 25 receiver. The transmitter has the function of processing the message

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signal into a form suitable for transmission over the channel. This processing of the message signal is referred to as modulation. The function of the channel is to provide a physical connection between the transmitter output and the receiver input. The function of the receiver is to process the received signal so as to produce an estimate of the original message signal. This processing of the received signal is referred to as demodulation.

One type of communication system is a spread-spectrum system. In a spread-spectrum system, a modulation technique is utilized in which a transmitted signal is spread over a wide frequency band within the communication channel. The frequency band is much wider than the minimum bandwidth required to transmit the information being sent. A voice signal, for example, can be sent with amplitude modulation (AM) in a bandwidth only twice that of the information itself. Other forms of modulation, such as low deviation frequency modulation (FM) or single sideband AM, also permit information to be transmitted in a bandwidth comparable to the bandwidth of the information itself. However, in a spread-spectrum system, the modulation of a signal to be transmitted often includes taking a baseband signal (e.g., a voice channel) with a bandwidth of only a few kilohertz, and distributing the signal to be transmitted over a frequency band that may be many megahertz wide. This is accomplished by modulating the signal to be transmitted with the information to be sent and with a wideband encoding signal.

Three general types of spread-spectrum communication techniques exist, including direct sequence modulation, frequency and/or time hopping modulation, and chirp modulation. In direct sequence modulation, a carrier signal is modulated by a digital code sequence whose bit rate is much higher than the information signal bandwidth.

Information (i.e., the message signal consisting of voice and/or data) can be embedded in the direct sequence spread-spectrum signal by several methods. One method is to add the information to the spreading code before it is used for spreading modulation. It will be noted that the information being sent must be in a digital form prior to adding it to the spreading code, because the combination of the spreading code and the information typically a binary code involves

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modulo-2 addition. Alternatively, the information or message signal may be used to modulate a carrier before spreading it.

These direct sequence spread-spectrum communication systems can readily be designed as multiple access communication systems.

5 For example, a spread-spectrum system may be designed as a direct sequence code division multiple access (DS-CDMA) system. In a DS-CDMA system, communication between two communication units is accomplished by spreading each transmitted signal over the frequency band of the communication channel with a unique user spreading code.
10 As a result, transmitted signals are in the same frequency band of the communication channel and are separated only by unique user spreading codes. These unique user spreading codes preferably are orthogonal to one another such that the cross-correlation between the spreading codes is low (i.e., approximately zero).

15 Particular transmitted signals can be retrieved from the communication channel by despreadng a signal representative of the sum of signals in the communication channel with a user spreading code related to the particular transmitted signal which is to be retrieved from the communication channel. Further, when the user spreading
20 codes are orthogonal to one another, the received signal can be correlated with a particular user spreading code such that only the desired user signal related to the particular spreading code is enhanced while the other signals for all of the other users are de-emphasized.

It will be appreciated by those skilled in the art that several
25 different spreading codes exist which can be used to separate data signals from one another in a DS-CDMA communication system. These spreading codes include but are not limited to pseudonoise (PN) codes and Walsh codes. A Walsh code corresponds to a single row or column of the Hadamard matrix.

30 Further it will be appreciated by those skilled in the art that spreading codes can be used to channel code data signals. The data signals are channel coded to improve performance of the communication system by enabling transmitted signals to better withstand the effects of various channel impairments, such as noise,
35 fading, and jamming. Typically, channel coding reduces the probability of bit error, and/or reduces the required signal to noise ratio (usually expressed as bit energy per noise density i.e., E_b/N_0 which is defined as

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the ratio of energy per information-bit to noise-spectral density), to recover the signal at the cost of expending more bandwidth than would otherwise be necessary to transmit the data signal. For example, Walsh codes can be used to channel code a data signal prior to modulation of the data signal for subsequent transmission. Similarly PN spreading codes can be used to channel code a data signal.

However, channel coding alone may not provide the required signal to noise ratio for some communication system designs which require the system to be able to handle a particular number of simultaneous communications (all having a minimum signal to noise ratio). This design constraint may be satisfied, in some instances, by designing the communication system to coherently detect transmitted signals rather than using non-coherent reception techniques. In coherent detection systems, the channel response is determined so that the affects of phase and magnitude distortions caused by the communication channel can be compensated for with matched filters. In contrast, non-coherent detection systems typically do not compensate for the phase distortion in a received signal which was caused by the communication channel. It will be appreciated by those skilled in the art that a coherent receiver requires less signal to noise ratio (in E_b/N_0) than that required by a non-coherent receiver having the same bit error rate (i.e., a particular design constraint denoting an acceptable interference level). Roughly speaking, there is a three deciBel (dB) difference between them for the Raleigh fading channel. The advantage of the coherent receiver is more significant when diversity reception is used, because there is no combining loss for an optimal coherent receiver while there is always a combining loss for noncoherent receiver.

One such method for facilitating coherent detection of transmitted signals is to use a pilot signal. For example, in a cellular communication system the forward channel, or down-link, (i.e., from base station to mobile unit) may be coherently detected if the base station transmits a pilot signal. Subsequently, all the mobile units use the pilot channel signal to estimate the channel phase and magnitude parameters. However, for the reverse channel, or up-link, (i.e., from mobile to base station), using such a common pilot signal is not feasible. As a result, those of ordinary skill in the art often assume that only non-coherent

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detection techniques are suitable for up-link communication. As a result, many recent publications have focused on optimizing non-coherent reception in DS-CDMA systems. Ideally a communication system should be designed to coherently received DS-CDMA signals.

5 However, channel coding and coherent detection still may not provide the required signal to noise ratio for some communication system designs which require the system to be able to handle a particular number of simultaneous communications (all having a minimum signal to noise ratio). It will be appreciated by those skilled in the art that in spread-spectrum CDMA communication systems the noise
10 interference level in the communication channel is directly related to number of simultaneous but code divided users within the communication channel. Thus, in order to reduce the noise interference, the number of simultaneous code divided users in the
15 communication channel is typically limited.

 In addition, the noise interference is also affected by the received signal power level. In some spread-spectrum communication systems (e.g., cellular systems) a central communication site typically attempts to detect or receive more than one signal from a particular band of the
20 electromagnetic frequency spectrum. The central communication site adjusts the receiver components to optimally receive signals at a particular received signal power threshold value. Those received signals having a received signal power level at or near the particular power threshold level are optimally received. While those received
25 signals not having a received signal power level at or near the particular power threshold level are not optimally received. A non-optimally received signal tends to have a higher bit error rate and tends to cause unnecessary interference to other receivers. Either of these undesirable consequences of a non-optimally received signal can result in the
30 communication system further limiting the number of simultaneous users in the communication channel associated with the central communication site.

 Thus, it is desirable to maintain the received signal power level at or near the particular power threshold level. This can be accomplished
35 by adjusting the signal power level of transmitters attempting to transmit to the central communication site. Therefore, by using power control schemes to maintain the received signal power levels at a particular

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power threshold level the number of simultaneous users in a communication channel can be maximized for a particular maximum error rate limit. This is especially important when vehicle speed is low. To achieve effective power control, an unbiased and accurate power estimate is needed. For non-coherent detection techniques in a DS-CDMA communication system, the use of orthogonal symbols, either bit interleaved or orthogonal symbol interleaved, is power estimation methods have been proposed. However, a need exists for a more accurate received signal power level estimation scheme for use in a coherent DS-CDMA spread-spectrum communication system. Through the use of a more accurate received signal power level estimation scheme, the number of simultaneous users in a communication channel can be increased over the number of simultaneous users in a communication channel using less accurate received signal power level estimation scheme while maintaining the same maximum error rate limit.

Summary of the Invention

A method and apparatus is provided for estimating signal power. This is accomplished by demodulating a received communication signal to derive a stream of samples. These samples may consist of either reference samples, data samples, or a combination thereof. An estimate of the power of the received communication signal is then generated as a function of the stream of samples. The signal power estimate may, for example, be used in a power control algorithm by setting a power control indicator in response to a comparison between the signal power estimate and a predetermined threshold. Subsequently, the power control indicator is transmitted over a communication channel to another communication unit. The communication unit detects a power control indicator within a signal received from over the communication channel and adjusts a particular signal transmission power of a signal transmitter in response to the detected power control indicator.

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Brief Description of the Drawings

FIG. 1 is a block diagram showing a preferred embodiment communication system in accordance with the present invention.

5 FIG. 2 is a block diagram showing a preferred embodiment communication channel frame structure for use in the preferred embodiment communication system shown in FIG. 1.

10 FIGs. 3 through 6 are block diagrams showing preferred embodiment signal power estimators for use in the preferred embodiment communication system shown in FIG. 1.

Detailed Description

15 In the course of the following discussion, a new approach for power control in up-link DS-CDMA communication systems is presented. This new approach is based on the coherent-detection with reference-symbol based channel estimation scheme which was described in a related patent application by Fuyun Ling entitled "Method and Apparatus for Coherent Communication in a Spread-Spectrum
20 Communication System" filed March 11, 1993 having U.S. Serial No. 08/031,258 and assigned to Motorola, Inc. In this coherent communication scheme, reference bits are uniformly inserted into convolutionally encoded bits and used for channel estimation. At the receiver side, the known reference bits are used to generate reference
25 symbols which have values of ± 1 . An accurate channel estimate is obtained by using these reference symbols and reference samples, which are the received signal samples corresponding to the reference bits, with a slight increased delay (e.g., approximately 4 milliseconds).

30 While such a short delay may not be a problem for the overall uplink communication, it cannot be tolerated for power estimation in a power control system. To reduce overall power control delay, only a few reference samples should be used to obtain a power estimate, which is an unbiased, but noise filled, estimate. Unfortunately, using a noise filled power estimate may reduce the effectiveness of power control.
35 This problem could be solved by using all of the received samples for power estimate through the maximum likelihood (ML) channel estimation method such as the one described in a related patent

application by David Falconer entitled "Path Gain Estimation in a Receiver" filed August 13, 1993 having U.S. Serial No. 08/106,250 and assigned to Motorola, Inc. A power estimate based on the ML method is indeed better than power estimate only based on the reference symbols. Nevertheless, the ML power estimate method is still not as effective as a power estimate based on orthogonal symbols used in the non-coherent DS-CDMA communication system design.

This inefficiency can be shown through simulations which indicate that the gain of coherent detection over non-coherent detection is greater at higher speeds than that at lower speeds. One such non-coherent DS-CDMA system is described in the communication standard known as IS-95 or "Mobile Station-Base Station Compatibility Standard for Dual-Mode Wideband Spread Spectrum Cellular System and published by the Electronic Industries Association (EIA), 2001 Eye Street, N.W., Washington, D.C. 20006. When using the ML power estimation scheme, the gain of a coherent detection scheme for a vehicle speed of 100 kilometers per hour (kph) is about 2.5 decibels (dB) over this non-coherent detection scheme (i.e., in IS-95). However, this gain is reduced to 1.7 dB for that of 15 kph. Thus, further improvement of power estimate is important, especially for low vehicle speeds. In the following description a power control method is revealed which uses both known reference symbols and received signal samples corresponding to the encoded bits that are not a priori known in estimating signal power. In order to put the power control method into an example environment, a basic coherent DS-CDMA communication system will be described. Subsequently, how the different aspects of power control can be applied to that basic communication system will be discussed.

It will be appreciated by those skilled in the art that, although the following discussion describes a cellular based communication system, other types of communication systems (e.g., personal communication systems, trunked systems, satellite communication systems, data networks, and the like) may also be adapted and/or designed to use the principles described herein.

Referring now to FIG. 1, a system for coherent communication in a spread-spectrum communication system is shown. In the encoding portion 100 of the communication system, traffic channel data bits 102

are input to an encoder 104 at a particular bit rate (e.g., 9.6 kilobit/second). The input traffic channel data bits can include either voice converted to data by a vocoder, pure data, or a combination of the two types of data. Encoder 104 encodes the input data bits 102 into data symbols at a fixed encoding rate (1/r) with an encoding algorithm which facilitates subsequent maximum likelihood decoding of the data symbols into data bits (e.g., convolutional or block coding algorithms). For example, encoder 104 encodes input data bits 102 (e.g., 192 input data bits that were received at a rate of 9.6 kilobits/second) at a fixed encoding rate of one data bit to three data symbols (i.e., 1/3) such that the encoder 104 outputs data symbols 106 (e.g., 576 data symbols output at a 28.8 kilo symbols/second rate).

The data symbols 106 are then input into an interleaver 108. Interleaver 108 organizes the data symbols 106 into blocks (i.e., frames) and block interleaves the input data symbols 106 at the symbol level. In the interleaver 108, the data symbols are individually input into a matrix which defines a predetermined size block of data symbols. The data symbols are input into locations within the matrix so that the matrix is filled in a column by column manner. The data symbols are individually output from locations within the matrix so that the matrix is emptied in a row by row manner. Typically, the matrix is a square matrix having a number of rows equal to the number of columns; however, other matrix forms can be chosen to increase the output interleaving distance between the consecutively input non-interleaved data symbols. The interleaved data symbols 110 are output by the interleaver 108 at the same data symbol rate that they were input (e.g., 28.8 kilo symbols/second). The predetermined size of the block of data symbols defined by the matrix is derived from the maximum number of data symbols which can be transmitted at a coded bit rate within a predetermined length transmission block. For example, if data symbols 106 are output from the encoder 104 at a 28.8 kilo symbols/second rate, and if the predetermined length of the transmission block is 20 milliseconds, then the predetermined size of the block of data symbols is 28.8 kilo symbols/second times 20 milliseconds (ms) which equals 576 data symbols which defines a 18 by 32 matrix.

The interleaved data symbols 110 are then input to a reference bit inserter 112 which inserts L known reference bits for every M

interleaved data symbols 110. To simplify the following discussion, it will be assumed that $L=1$ and $M=6$. In addition, it will be assumed that each inserted reference bit is a zero bit. However, it will be appreciated by those skilled in the art that L and M could be any other value without departing from the scope and spirit of the present invention. In addition, the reference bits could be any known sequence such as all one bits or several one bits followed by several zero bits without departing from the scope and spirit of the present invention. When $L=1$ and $M=6$, reference bit inserter 112 outputs 672 reference-coded symbols 114 for each block (i.e., frame) such that a reference bit is inserted between each group of six data symbols. An example of a transmitted block (i.e., frame) of reference-coded data symbols 114 consisting of 42 bits is shown in FIG. 2 (where each d represents a data symbol and each r represents a reference bit).

The reference-coded data symbols 114 is output from encoding portion 100 of the communication system and input to a transmitting portion 116 of the communication system. The data symbols 114 are prepared for transmission over a communication channel by a modulator 117. Subsequently, the modulated signal is provided to an antenna 118 for transmission over the communication channel 120.

The modulator 117 preferably prepares the data symbols 114 for direct sequence code divided spread-spectrum transmission by deriving a sequence of fixed length codes from the reference-coded data symbols 114 in a spreading process. For example, each of the data symbols within the stream of reference-coded data symbols 114 may be spread to a unique nine bit length code such that a group of six data symbols is represented by a single 54 bit length code. In addition, each reference bit within the stream of reference-coded data symbols 114 may select a ten bit length code. The codes representing the group of six data symbols and an associated reference bit preferably are combined to form a single 64 bit length code. As a result of this spreading process, the modulator 117 which received the reference-coded data symbols 114 at a fixed rate (e.g., 28.8 kilo symbols/second) now has a spread sequence of 64 bit length codes having a higher fixed symbol rate (e.g., 307.2 kilo symbols/second). It will be appreciated by those skilled in the art that the reference symbols and data symbols within the stream of reference-coded data bits 114 may be spread

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according to numerous other algorithms into a sequence of larger length codes without departing from the scope and spirit of the present invention.

5 The spread sequence is further prepared for direct sequence code divided spread-spectrum transmission by further spreading the spread sequence with a long spreading code (e.g., PN code). The spreading code is a user specific sequence of symbols or unique user code which is output at a fixed chip rate (e.g., 1.228 Megachips/second). In addition to providing an identification as to which user sent the
10 encoded traffic channel data bits 102 over the communication channel 120, the unique user code enhances the security of the communication in the communication channel by scrambling the encoded traffic channel data bits 102. In addition, the user code spread encoded data bits (i.e., data symbols) are used to bi-phase modulate a sinusoid by
15 driving the phase controls of the sinusoid. The sinusoid output signal is bandpass filtered, translated to an RF frequency, amplified, filtered and radiated by an antenna 118 to complete transmission of the traffic channel data bits 102 in a communication channel 120 with BPSK modulation.

20 A receiving portion 122 of the communication system receives the transmitted spread-spectrum signal from over the communication channel 120 through antenna 124. The received signal is sampled into data samples by despreaders and sampler 126. Subsequently, the data samples 142 are output to the decoding portion 154 of the
25 communication system.

The despreaders and sampler 126 preferably BPSK samples the received spread-spectrum signal by filtering, demodulating, translating from the RF frequencies, and sampling at a predetermined rate (e.g., 1.2288 Megasamples/second). Subsequently, the BPSK sampled
30 signal is despread by correlating the received sampled signals with the long spreading code. The resulting despread sampled signal 128 is sampled at a predetermined rate and output to a reference bit (sample) extractor 130 (e.g., 307.2 kilo samples/second so that a sequence of four samples of the received spread-spectrum signal is despread and/or
35 represented by a single data sample).

The reference bit extractor 130 preferably extracts the reference samples 132 from the despread sampled signal 128 and outputs the

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reference samples 132 to a channel estimator 134. The remaining data samples 138 from the despread sampled signal 128 are output to a coherent detector 140 for later coherent detection of data samples 142.

5 The channel estimator 134 correlates the extracted reference samples 132 with a known reference sequence of reference symbols to obtain unbiased, but noisy, channel estimates. In order to obtain a better channel estimate 136, these noisy estimates may be passed through a low-pass filter, which may be fixed or adaptive, to remove the
10 high frequency noise components. The resulting channel estimates 136 are relatively noise-free and can be used for coherent detection. It should be noted that the low pass filtering only gives us a channel estimate for every $(M+1)T$, where M is the number of data symbols between each reference bit inserted by reference bit inserter 112 (e.g.,
15 $M=6$) and where T is the time interval of each data sample. To perform coherent detection of transmitted data symbols, we need to have a channel estimate for every T . When $(M+1)T$ is short relative to the channel variation time constant, a simple but effective method to get a channel estimate for every T is to perform linear interpolation between
20 two channel estimates separated by $(M+1)T$. However, as will be appreciated by those skilled in the art more sophisticated interpolation techniques may be used if necessary.

After channel estimates 136 are generated, the rest of the receiver is conventional. The coherent detector 140 multiplies the
25 remaining data samples 138 from the despread sampled signal 128 by the conjugate of the channel estimates 136 to generate coherently detected samples 142.

As will be appreciated by those skilled in the art, multiple receiving portions 122 through 123 and antennae 124 through 125,
30 respectively, can be used to achieve space diversity. The N th receiver portion would operate in substantially the same manner to retrieve data samples from the received spread-spectrum signal in communication channel 120 as the above described receiving portion 122. The outputs 142 through 152 of the N receiving portions preferably are
35 input to a summer 150 which diversity combines the input data samples into a composite stream of coherently detected data samples 160.

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The individual data samples **160** which form soft decision data are then input into a decoding portion **154** including a deinterleaver **162** which deinterleaves the input soft decision data **160** at the individual data level. In the deinterleaver **162**, the soft decision data **160** are individually input into a matrix which defines a predetermined size block of soft decision data. The soft decision data are input into locations within the matrix so that the matrix is filled in a row by row manner. The deinterleaved soft decision data **164** are individually output from locations within the matrix so that the matrix is emptied in a column by column manner. The deinterleaved soft decision data **164** are output by the deinterleaver **162** at the same rate that they were input (e.g., 28.8 kilometrics/second).

The predetermined size of the block of soft decision data defined by the matrix is derived from the maximum rate of sampling data samples from the spread-spectrum signal received within the predetermined length transmission block.

The deinterleaved soft decision data **164**, are input to a decoder **166** which uses maximum likelihood decoding techniques to generate estimated traffic channel data bits **168**. The maximum likelihood decoding techniques may be augmented by using an algorithm which is substantially similar to a Viterbi decoding algorithm. The decoder **166** uses a group of the individual soft decision data **164** to form a set of soft decision transition metrics for use at each particular time state of the maximum likelihood sequence estimation decoder **166**. The number of soft decision data **164** in the group used to form each set of soft decision transition metrics corresponds to the number of data symbols **106** at the output of the convolutional encoder **104** generated from each input data bit **102**. The number of soft decision transition metrics in each set is equal to two raised to the power of the number of soft decision data **164** in each group. For example, when a 1/3 convolutional encoder is used in the transmitter, three data symbols **106** are generated from each input data bit **102**. Thus, decoder **166** uses groups of three individual soft decision data **164** to form eight soft decision transition metrics for use at each time state in the maximum likelihood sequence estimation decoder **166**. The estimated data bits **168** are generated at a rate related to the rate that the soft decision data **164** are input to the decoder **166** and the fixed rate used to originally

encode the input data bits **102** (e.g., if the soft decision data are input at 28.8 kilometrics/second and the original encoding rate was 1/3 then estimated data bits **168** are output at a rate of 9600 bits/second).

Thus, a communication system for coherently encoding and
5 decoding has been described above with reference to FIG. 1. In summary, the communication system includes a first portion which encodes input data bits into data symbols, interleaves the data symbols in a symbol by symbol manner, inserts reference bits into the interleaved symbols, modulates and transmits the reference-coded data symbols
10 over a communication channel. The communication system further includes a second portion which receives and demodulates a signal from over the communication channel, estimates parameters of the communication channel, coherently demodulates data samples within received signal, deinterleaves the coherently detected data samples
15 which are used as soft decision data within each received transmission block, subsequently generates soft decision transition metrics from the deinterleaved individual soft decision data, and subsequently generates estimated data bits from the soft decision metrics by using maximum likelihood decoding techniques.

20 Turning now to the power control scheme for use in the preferred embodiment coherent communication system, the preferred embodiment power control algorithm preferably includes estimating received power every 1.25 ms (i.e., each block or frame) like the DS-CDMA communication system described in IS-95, or every 12
25 information bits, i.e., every 36 encoded data symbols or 42 total received signal samples. The power control algorithm includes despreading **126** a communication signal received **124** from over a communication channel **120** with a spreading code to derive a stream of samples, at one communication unit. This stream of samples may consist of
30 reference samples, data samples, or a combination of both. The stream of samples **144** is input to a power estimator **146** so that an estimate **148** of the power of the received communication signal as a function of the stream of samples can be generated. It will be appreciated by those skilled in the art that in a diversity combining receiver, it is desirable to
35 generate a combined power estimate from all of the diversity paths. Subsequently, a comparator or threshold device **169** sets a power control indicator **171** in response to the result of a comparison between

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the signal power estimate 148 and a predetermined threshold. This power control indicator 171 is prepared for transmission and transmitted through antenna 174 over the communication channel 120 by a modulator 170. It will be appreciated that other control information 5 172 (e.g., user authentication and channel hand-off information) may also be prepared and transmitted by modulator 170. In a DS-CDMA communication system, the power control indicator 171 may be prepared for transmission over the communication channel 120 by spreading the power control indicator 171 with a spreading code prior 10 to transmission over the communication channel 120.

This power control indicator is then received by another communication unit. The power control indicator is received within a signal from over the communication channel 120 by antenna 176. Like the transmission process, for a DS-CDMA communication system, the 15 power control indicator 171 may need to be despread 178 from the received signal with a spreading code prior to complete detection of the power control indicator 171. Once the power control indicator 180 is detected, the communication unit may adjust in the modulator 117 a particular signal transmitter power of the signal transmitter 116 in 20 response to the detected power control indicator 180. In addition, the other detected control information 182 (e.g., user authentication and channel hand-off information) can be used by the receiving communication unit.

The main part of this power control scheme is the acquisition of a 25 power estimate 148. In this coherent communication system, the despreaders 126 outputs N received samples $r_s(k)$, corresponding to the encoded symbols (i.e., signal samples) and M samples $r_{ref}(k)$, corresponding to the reference bits (i.e., reference samples) for each power estimation period. By way of example for the following 30 discussion, $N = 36$ and $M = 7$. However, it will be appreciated by those skilled in the art that N and M may be varied to suit the needs of a particular communication system without departing from the scope and spirit of the present invention. Several power estimation algorithms could be used in this power control scheme.

35 In a DS-CDMA uplink communication system, the received signal samples $r(n)$ 144 after despreading 126 can be expressed in the form:

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$$r(n) = a(n)c(n) + z(n) \quad (\text{Eq. 1})$$

where $a(n)$ represents either the convolutionally coded symbols mapped into values of ± 1 or the reference symbols, $c(n)$ is channel coefficient which is a complex number, and $z(n)$ is interference term which mainly consists of the signal from other users in the communication system. It shall be assumed that the power estimation period is 1.25 ms, like in IS-95, which is relatively short in terms of communication channel variations. Therefore, the communication channel can be considered unvarying or constant during each power estimation period and expressed simply as c .

It will be appreciated by those skilled in the art that the received signal power (s) during an estimation period is simply:

$$s = E[|a(n)c|^2] = |c|^2 \quad (\text{Eq. 2})$$

because $|a(n)|^2 = 1$.

When, in a first power estimation scheme, $a(n)$ is known (i.e., for reference symbols $a_r(n)$), the received signal power from (Eq. 1) can be expressed as:

$$s_{\text{ref}} = \frac{1}{7} \sum_{k=0}^6 [|a_r(n)|^2 c + a_r^*(n) z(n)] = c + \frac{1}{7} \sum_{k=0}^6 a_r^*(n) z(n), \quad (\text{Eq. 3})$$

where as previously noted $|a_r(n)|^2 = 1$. Thus, it will be appreciated that the sum on the right hand side of the (Eq. 3) is a Gaussian zero mean variable with a variance which is 1/7 of the variance of $z(n)$. Which means that s_{ref} is an unbiased estimate of c . By writing c as $|c|e^{j\theta_c}$, the absolute value of s_{ref} from (Eq. 3) can be written as:

$$|s_{\text{ref}}| = \left| c + \frac{1}{7} \sum_{k=0}^6 a_r^*(n) z(n) \right| = |c| \left| 1 + \frac{1}{|c|} e^{-j\theta_c} \frac{1}{7} \sum_{k=0}^6 a_r^*(n) z(n) \right| \quad (\text{Eq. 4})$$

If the signal noise variance is not large relative to $|c|$, then

$1 + \frac{1}{|c|} e^{-j\theta_c} \frac{1}{7} \sum_{k=0}^6 a_r^*(n) z(n)$ is unlikely become negative. In such a case, $|s_{\text{ref}}|$ is a good estimate of $|c|$.

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In the power control approach for IS-95, it is only necessary to determine if the received signal power (i.e., $|c|^2$) is larger or smaller than a predetermined threshold. This is equivalent to determining if $|c|$ is larger or smaller than the square-root of the original threshold. Thus, for a coherent DS-CDMA communication system which uses a similar power control approach, $|s_{ref}|$ can be directly used for power control.

In a second alternative power estimation scheme, $a(n)$ consists of mapped convolutionally coded symbols $a_s(n)$. For this scheme, the received signal power from (Eq. 1) can be expressed as:

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$$\begin{aligned} s_s &= \frac{1}{36} \sum_{k=0}^{35} r_s^2(k) = \frac{1}{36} \sum_{n=0}^{35} (a(n)c + z(n))^2 \\ &= \frac{1}{36} \sum_{n=0}^{35} (a_s^2(n)c^2 + 2a_s^2(n)cz(n) + z^2(n)) \\ &= c^2 + \frac{1}{36} \sum_{n=0}^{35} (2a_s^2(n)cz(n) + z^2(n)) \equiv c^2 + v \end{aligned} \quad (\text{Eq. 5})$$

where v is approximately a zero-mean Gaussian random variable.

Thus, s_s is an unbiased estimate of c^2 . Furthermore, the variance of v is small due to averaging.

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By writing c^2 as:

$$c^2 = |c|^2 e^{j2\phi_c} \quad (\text{Eq. 6})$$

From (Eq. 5) and (Eq. 6) we can show that $|s_s|$ can be expressed as:

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$$|s_s| = |c|^2 + v e^{-j2\phi_c} = |c|^2 \left| 1 + \frac{1}{|c|^2} v e^{-j2\phi_c} \right| \quad (\text{Eq. 7})$$

Using the same argument as above, if the variance of v is not too large relative to $|c|^2$, $|s_s|$ is a good estimate of $|c|^2$. Thus we can use $|s_s|$ as the power estimate for power control.

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In a third alternative power estimation scheme, $a(n)$ consists of mapped convolutionally coded symbols $a_s(n)$ and reference symbols $a_r(n)$. Further, it can be shown that s_{ref}^2 is an unbiased estimate of c^2 .

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Thus, s_{ref}^2 and s_s may be combined to obtain a better estimate of c^2 . Similarly, the square-root of s_s is an estimate of $|c|$, so that the square-

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root of s_s and s_{ref} can be combined to obtain a better estimate of $|c|$. These better estimates of $|c|$ and c^2 can be used as power estimates in a power control mechanism.

Referring now to the first power estimation algorithm, shown in FIG. 3, only reference samples 144 are used for power estimation 146. In using only the reference samples 144 for generating 146 a power estimate 148, only a scaled sum of a product needs to be calculated. For example, in (Eq. 8) which is derived from (Eq. 4), a preferred embodiment reference sample based is shown.

$$s_{ref} = \frac{1}{7} \sum_{k=0}^6 a_r^*(k) r_{ref}(k) \quad (\text{Eq. 8})$$

where $a_r^*(k)$, $k = 0$ to 6 , are the reference symbols. As has been discussed above, the magnitude, or absolute value, of s_{ref} (i.e., $|s_{ref}|$) can be used for power control as an approximation of a power estimate, although it is actually not a power estimate. This is implemented as summing device 200 which sums the stream of seven reference symbols (i.e., samples) together and inputs the sum 202 into a combiner 206. Combiner 206 multiplies the summed reference samples with a scale factor 204 (e.g., a $1/7$ scale factor). This summing and multiplying operation results in a complex signal power estimate 208 (i.e., an estimate which has a real and imaginary part). This complex signal power estimate 208 is transformed into a reference-based signal power estimate 148 by calculating 210 the magnitude of the average complex reference signal power estimate 208. It will be appreciated by those skilled in the art that the long term accuracy of this complex signal power estimate can be improved by averaging several estimates together to form an average complex reference signal power estimate.

Referring now to the second alternative power estimation algorithm shown in FIG. 4, only convolutionally coded samples 144 are used for power estimation 146. A square of the received signal samples are determined and a scaled sum of these squared values is formed, as derived from (Eq. 5) as follows:

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$$s_s = \frac{1}{36} \sum_{k=0}^{35} r_s^2(k) \quad (\text{Eq. 9})$$

Note that $r_s(k)$ are complex numbers and s_s is also a complex value. As has been previously shown, the magnitude of s_s , i.e., $|s_s|$, is an estimate of the signal power and can be used for power control. This is implemented with a complex squaring device 212 which squares convolutionally coded data samples 144. These thirty-six complex squared data samples 214 are then summed together by a summer 216. Subsequently, the summed complex data samples 218 are transformed into a data-based signal power estimate 222 by calculating its magnitude. This data-based signal power estimate 222 is then properly scaled by a combiner 226 which multiplies it with a scale factor 224 (e.g., a 1/36 scale factor). This properly scaled data-based signal power estimate 148 is then output by power estimator 146 for use in a power control algorithm. It will be appreciated by those skilled in the art that, like the first power estimate, the long term accuracy of this signal power estimate can be improved by averaging several complex squared data samples together to form an average complex squared data sample prior to generating the data-based signal power estimate 148 from a magnitude operation 220.

Referring now to the third alternative power estimation algorithm, shown in FIGs. 5 and 6, both reference samples and convolutionally coded data samples 144 are used for power estimation 146. More accuracy may be achieved by using both the reference and signal samples for power estimation 146. There are a few different ways to combine these samples. The first method is to simply using a variation of (Eq. 9) by treating $r_{\text{ref}}(k)$ no different from $r_s(k)$ and forming the power estimate p_1 given by:

$$p_1 = \left| \frac{1}{43} \sum_{k=0}^{42} r^2(k) \right| \quad (\text{Eq. 10})$$

where $r(k)$ represents both $r_{\text{ref}}(k)$ and $r_s(k)$. Because more samples are used, the estimate given by (Eq. 10) is better than that given by (Eq. 9).

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The second method (shown in FIG. 5) is to form a sum 258 of s_s and the square value of s_{ref} (which are performed by elements 228-240 for s_{ref} and elements 244-256 for s_s). The magnitude 260 of the sum 258 is used as the power estimate 148. To obtain a better unbiased estimate, both values 240 and 256 should be weighted as a linear combination before forming 242 the sum 258. More precisely, the power estimate p_2 is formed by:

$$p_2 = \left| \frac{2}{3} s_s + \frac{1}{3} s_{ref}^2 \right|. \quad (\text{Eq. 11})$$

10

The third method (shown in FIG. 6) is to form a weighted sum 148 of the absolute value of the real part of s_{ref} and the square root of the magnitude of s_s (which are performed by elements 262-274 for s_{ref} and elements 278-294 for s_s). Similar to (Eq. 11) in order to obtain a better unbiased estimate, values 274 and 294 should be weighted as a linear combination before forming 276 the sum 148. More precisely, the power estimate p_3 is formed by:

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$$p_3 = \frac{2}{3} |s_{ref}| + \frac{1}{3} \sqrt{|s_s|} \quad (\text{Eq. 12})$$

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It will be appreciated by those skilled in the art that the weighting factors may be modified for a particular communication system without departing from the scope or spirit of the present invention.

The second and third methods would give similar results. However, the second method is preferable because it is computationally simpler. It is also much simpler than the previously mentioned ML power estimation method. It is expected that the performance of the coherent DS-CDMA receiver can be improved at 15 kph vehicle speeds by about 0.7 dB over the ML power estimating scheme. Further improvements are also possible for very low vehicle speeds. As a result, the coherent DS-CDMA receiver using one of the above-mentioned signal power estimates for power control should consistently outperform the IS-95 based DS-CDMA receiver by about 2.2 dB without diversity combining and about 2.6 dB with two port diversity combining.

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Although the invention has been described and illustrated with a certain degree of particularity, it is understood that the present disclosure of embodiments has been made by way of example only and that numerous changes in the arrangement and combination of parts as well as steps may be resorted to by those skilled in the art without departing from the spirit and scope of the invention as claimed. For example, the modulator, antennas and demodulator portions of the preferred embodiment communication system as described were directed to CDMA spread-spectrum signals transmitted over a radio communication channel. However, as will be understood by those skilled in the art, the encoding and decoding techniques described and claimed herein can also be adapted for use in other types of transmission systems like those based on time division multiple access (TDMA) and frequency division multiple access (FDMA). In addition, the communication channel could alternatively be an electronic data bus, wireline, optical fiber link, satellite link, or any other type of communication channel.

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Claims

What is claimed is:

- 5 1. An apparatus comprising a signal power estimator, the signal power estimator comprising:
 - (a) demodulating means for demodulating a received communication signal to derive a stream of samples selected from the group consisting of: a stream of
10 reference samples and a stream of data samples; and
 - (b) estimating means, operatively coupled to the demodulating means, for generating an estimate of the power of the received communication signal as a function of the stream of samples.
- 15 2. The apparatus of claim 1 wherein the estimating means comprises means for generating a plurality of complex signal power estimates from the stream of reference samples by correlating a known reference sequence with the stream of
20 reference samples.
3. The apparatus of claim 2 wherein the estimating means further comprises averaging means for averaging the plurality of complex signal power estimates together to form an average
25 complex reference signal power estimate.
4. The apparatus of claim 3 wherein the estimating means further comprises means for generating a reference-based signal power estimate by calculating the magnitude of the average complex
30 reference signal power estimate.
5. The apparatus of claim 1 wherein the estimating means comprises means for generating a plurality of complex squared data samples from the stream of data samples by complex
35 squaring the stream of data samples.

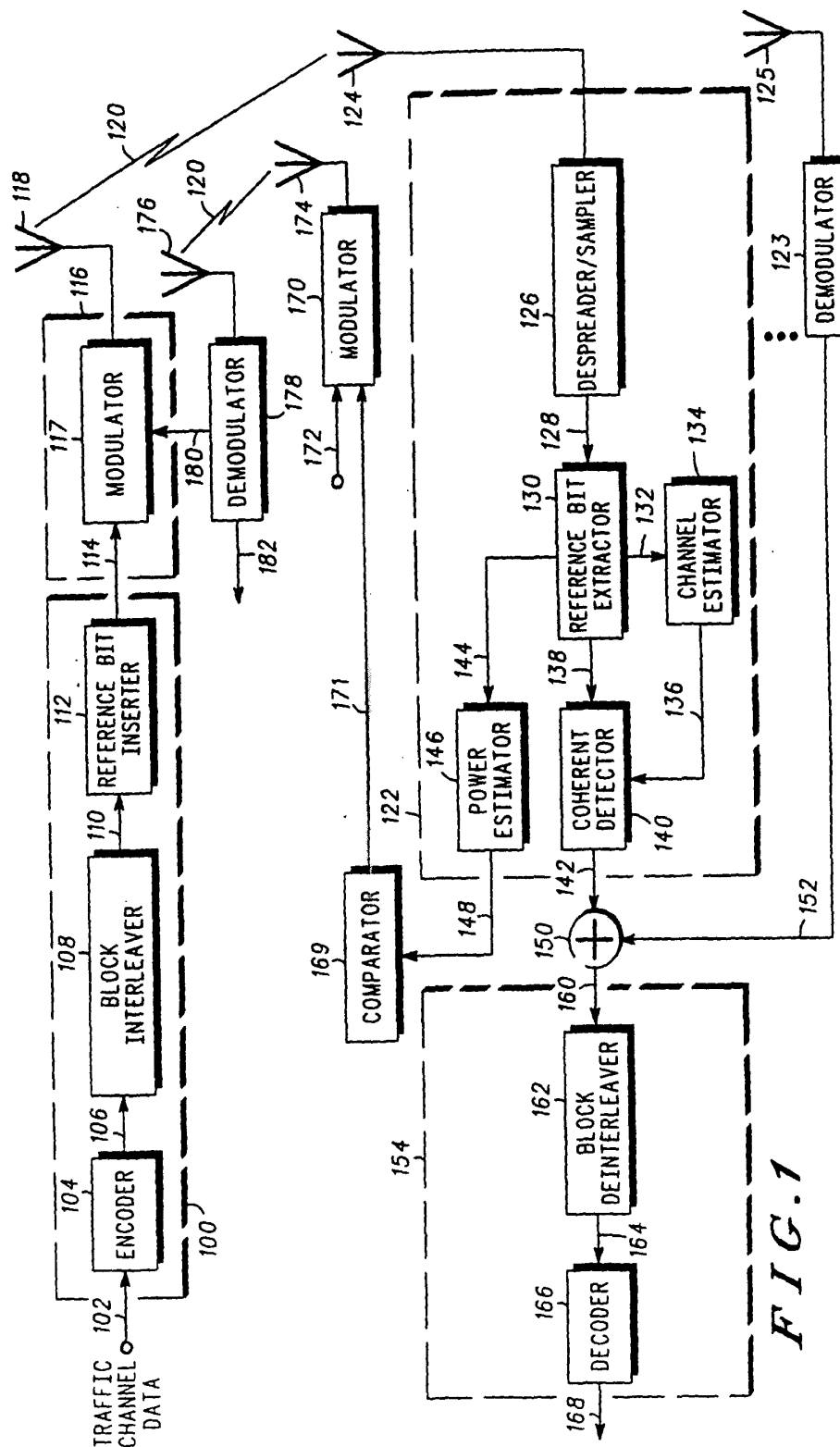
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6. The apparatus of claim 5 wherein the estimating means further comprises averaging means for averaging the plurality of complex squared data samples together to form an average complex squared data sample.
- 5
7. The apparatus of claim 6 wherein the estimating means further comprises means for generating a data-based signal power estimate by calculating the magnitude of the average complex squared data sample.
- 10
8. A communication unit, comprising:
- (a) signal receiving means for detecting a power control indicator within a signal received from over the communication channel, the power control indicator being derived from a comparison between a signal power estimate and a predetermined threshold, the signal power estimate being derived from a despread communication signal as a function of a stream of samples selected from the group consisting of: a stream of reference samples and a stream of data samples;
- 15
- 20
- (b) power adjustment means, operatively coupled to the signal receiving means, for adjusting a particular signal transmission power of a signal transmitter in response to the detected power control indicator.
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9. The apparatus of claim 8 wherein the signal receiving means comprises means for despread the received signal with a spreading code to detect the power control indicator .

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10. A method of estimating signal power, comprising:
- 5 (a) demodulating a received communication signal to derive a stream of samples selected from the group consisting of: a stream of reference samples and a stream of data samples; and
- (b) generating an estimate of the power of the received communication signal as a function of the stream of samples.

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FIG. 5

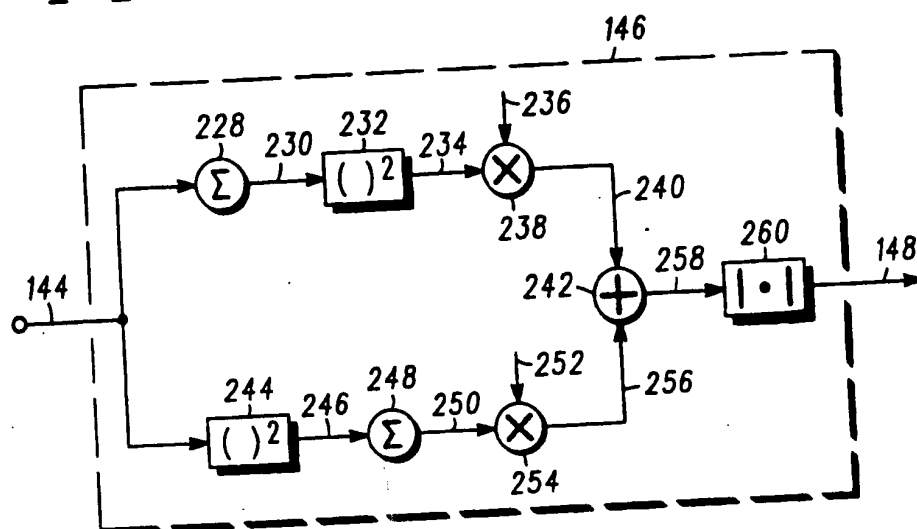
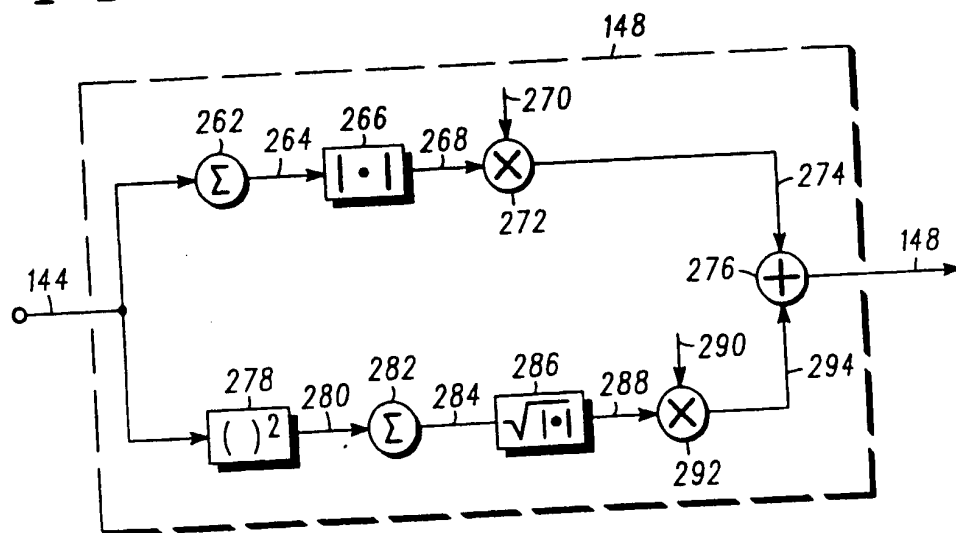


FIG. 6



INTERNATIONAL SEARCH REPORT

International application No.
PCT/US94/08472

A. CLASSIFICATION OF SUBJECT MATTER

IPC(5) :H04B 7/216

US CL :375/1

According to International Patent Classification (IPC) or to both national classification and IPC

B. FIELDS SEARCHED

Minimum documentation searched (classification system followed by classification symbols)

U.S. 375/1; 342/63; 455/67.1; 370/32.1

Documentation searched other than minimum documentation to the extent that such documents are included in the fields searched

Electronic data base consulted during the international search (name of data base and, where practicable, search terms used)

C. DOCUMENTS CONSIDERED TO BE RELEVANT

Category*	Citation of document, with indication, where appropriate, of the relevant passages	Relevant to claim No.
Y	US, A, 4,591,669 (DUTTWEILER ET AL) 27 MAY 1986. See Figures 1-6.	1-10
Y	US, A, 5,216,692 (LING) 01 JUNE 1993. See Figures 1-3.	1-10
A,P	US, A, 5,297,161 (LING) 22 MARCH 1994. See Figures 1-3.	1-10
A,P	US, A, 5,329,547 (LING) 12 JULY 1994. See Figures 1-3.	1-10
Y	IEEE TRANSACTIONS ON INFORMATION THEORY, VOL. IT-29, No. 4, JULY 1983, VITERBI ET AL, "NONLINEAR ESTIMATION OF PSK-MODULATED CARRIER PHASE WITH APPLICATION TO BURST DIGITAL TRANSMISSION, SEE PAGES 543-551.	1-10

☐ Further documents are listed in the continuation of Box C. ☐ See page

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"1" document referring to an oral disclosure, use, exhibition or other means		
"1" document published prior to the international filing date but later than the priority date claimed	"&"	document member of the same patent family

Date of the actual completion of the international search

01 SEPTEMBER 1994

Date of mailing of the international search report

22 SEP 1994

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